A New Approach Fuzzy-MRAS-Sensorless vector Control Based Rotor and Stator Resistances Estimation of an Induction Motor Drive

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Abstract: *This paper introduces a new sensorless vector controlled induction motor based to the parallel resistance parameters and rotor speed estimation for Fuzzy-MRAS Sensorless with Direct Field Oriented Control is discussed in this paper. the analysis of the sensorless vector control system using MRAS is presented and the resistance parameters variations and speed observer using Fuzzy Logic is proposed. The aim of the proposed sensorless control is to improve the performance and robustness of the induction motor drives under non linear loads variations is presented in this work. The availability of the proposed structure scheme is tested in simulation with Matlab-software. The Simulation results will be presented to confirm the theoretical analysis and the efficiency and reliability of the proposed Fuzzy-MRAS sensorless.*

Index Terms-- Induction motor drives, vector control, fuzzy logic, fuzzy-MRAS observer, robustness.

1. Nomenclature

s, r : Stator and rotor subscripts
d, q : Direct and quadrate Park subscripts
v, i, φ : Voltage/ Current/ Flux variables
R_S, R_R : Stator, rotor resistances
L_S, L_R : Stator, rotor inductance
L_m : Mutual magnetizing inductances
σ : Total leakage factor
ω_S : Stator frequency
ω_r : Slip frequency
ω_m : Nominal frequency
Ω : Rotor speed
θ_s : Rotor flux position
J : Inertia
f : Friction coefficient
T_e : Electromagnetic Torque
p : Pole pair number
^ : Superscript of estimated quantity
IM : Induction Motor

2. Introduction

Recently, the sensorless control of induction motor drives constitutes a vast subject, and the technology has further advanced in recent years. The control of the asynchronous machine, however, is complex because the dynamics of the machine are non-linear, multivariable, and highly coupled. Furthermore, there are various uncertainties and disturbances in the system. In high performance applications, the induction motor is controlled through field orientation technique. The basic goal of the FOC is to resolve the stator current vector into two components: one used to control the machine flux and the other to control the machine torque, thus, enabling the torque and the flux to be controlled independently [1],[2]. The Sensorless induction motor control based upon the theory of Model Reference Adaptive System (MRAS) provides an alternative way for the development. The MRAS speed estimators are most attractive due to their design simplicities, in which the outputs of to models, one independent of the rotor speed (reference model) and the other dependent (adjustable model), are used to form an error vector. The error vector is driven to zero by an adaptation mechanism, which yields the estimated rotor speed. However, this method requires both rotor and stator resistors, and integral operation which involves initial value problem. Although, stator resistance and rotor time constant which are sensitive to temperature variation, a model reference adaptive system technique is proposed to estimate the stator resistance, rotor resistance and rotor speed estimation by means of MRAS observers [3],[4]. The rotor speed for adaptive mechanism is estimated by fuzzy logic controllers. The fuzzy logic have gained great important and witnessed a rapid growth in industrial applications, and proved their dexterity of many respects [5],[6]. They proved that such control can achieve satisfactory results in dealing with systems whose behaviour is difficult to describe mathematically or is highly nonlinear. The paper is organized as follows: the model dynamic of induction motor is presented in Section 3, after the direct field oriented control is presented in Section 4. In section 5, a fuzzy logic controller is presented. In section 6 the structure of the proposed Fuzzy-MRAS with resistance parameters and rotor speed estimation is described. In section 7, and through simulation, the studied estimated Fuzzy-MRAS is associated to the direct-field-oriented control where stator resistance, rotor resistance and rotor speed was replaced by those delivered by the estimated. Finally, in section 8, we give some comments and conclusions.

3. DIFFERENTIAL EQUATIONS OF IM

Mathematical description of the induction motor is based on complex space vectors, which are defined in a coordinate system rotating with angular speed. In per unit and real time representation the following vector equations describe behaviour of the motor. The dynamics of the induction

motor in the *d-q* motor reference frame, which is rotating at the synchronously speed, can be simply described by the following nonlinear differential [1]:

$$\begin{cases} \dfrac{d}{dt} i_{sd} = \dfrac{1}{\sigma \cdot L_s} \Big[-\big(R_s + \dfrac{L_m^2}{L_r \cdot T_r}\big) \cdot i_{sd} + \omega_s \cdot \sigma \cdot L_s \cdot i_{sq} \\ \qquad + \dfrac{L_m}{L_r \cdot T_r} \cdot \phi_{rd} + \dfrac{L_m}{L_r} \cdot \omega_r \cdot \phi_{rq} + v_{sd} \Big] \\ \dfrac{d}{dt} i_{sq} = \dfrac{1}{\sigma \cdot L_s} \Big[-\omega_s \cdot \sigma \cdot L_s \cdot i_{sd} - \left(R_s + \dfrac{L_m^2}{L_r \cdot T_r}\right) \cdot i_{sq} \\ \qquad - \dfrac{L_m}{L_r} \omega_r \cdot \phi_{rd} + \dfrac{L_m}{L_r \cdot T_r} \cdot \phi_{rq} + v_{sq} \Big] \\ \dfrac{d}{dt} \phi_{rd} = \dfrac{L_m}{T_r} \cdot i_{sd} - \dfrac{1}{T_r} \cdot \phi_{rd} + (\omega_s - \omega_r) \cdot \phi_{rq} \\ \dfrac{d}{dt} \phi_{rq} = \dfrac{L_m}{T_r} \cdot i_{sq} - (\omega_s - \omega_r) \cdot \phi_{rd} - \dfrac{1}{T_r} \cdot \phi_{rq} \\ \dfrac{d}{dt} \omega = \dfrac{P^2 \cdot L_m}{L_r \cdot J} (i_{sq} \phi_{rd} - i_{sd} \phi_{rq}) - \dfrac{F}{J} \omega - \dfrac{P}{J} C_r \\ \qquad \text{where } \sigma = 1 - \dfrac{L_m^2}{L_s L_r} \end{cases} \tag{1}$$

4. Direct Field Oriented Control

Direct Field oriented control (DFOC) technique is intended to control the motor flux, and thereby be able to decompose the AC motor current into "flux producing" and "torque producing" components. The well known direct field orientation strategies provide a linear and decoupled control between the flux and torque of the induction machine [7],[8]. Then the rotor flux orientation process is given by the imposed zero constraint of quadrate rotor flux component. Such as:

$\phi_{rq}=0$ and $\phi_{rd}=\phi_r$ (2)

Hence, the rotor flux can be controlled directly from the stator direct current component i_{sd}, while the torque can be linearly controlled from the stator quadrate current component i_{sq} when the rotor flux is maintained constant. Separating the real and imaginary parts of (1) by using (2) leads to:

$$\begin{cases} v_{sd} = \sigma L_s \dfrac{d}{dt} i_{sd} + \left(R_s + R_r \dfrac{L_m^2}{L_r^2} \right) i_{sd} - \omega_s \sigma L_s i_{sq} - \dfrac{L_m}{L_r^2} R_r \phi_r \\ \quad = v_{sd}^r - \omega_s \sigma L_s i_{sq} \\ v_{sq} = \sigma L_s \dfrac{d}{dt} i_{sq} + \omega_s \sigma L_s i_{sd} + \left(R_s + R_r \dfrac{L_m^2}{L_r^2} \right) i_{sq} + \dfrac{L_m}{L_r^2} p \Omega \phi_r \\ \quad = v_{sq}^r - \omega_s \sigma L_s i_{sd} + \dfrac{L_m}{L_r^2} p \Omega \phi_r \end{cases} \tag{3}$$

The slip frequency can be calculated from the values of the stator current quadrate and the rotor flux oriented reference frame as follow:

$$\omega = \omega_s - \omega_r = \frac{L_m}{T_r} \frac{i_{sq}}{\phi_r} \tag{4}$$

And the rotor flux position is given by:

$$\theta_s = \int \omega_s dt \tag{5}$$

The voltages v_{sd} and v_{sq} should act on the current i_{sd} and i_{sq} separately and consequently the flux and the torque. The two-phase stators current are controlled by two PI controllers taking as input the reference values i_{sd}^*, i_{sq}^* and the measured values. Thus, the common thought is to realize the decoupling by adding the compensation terms (e_{sd} and e_{sq}) as usually done [11].

$$\begin{cases} e_{sd} = \omega_s \sigma L_s i_{sq} \\ e_{sq} = \omega_s \sigma L_s i_{sd} - \omega_s \dfrac{L_m}{L_r^2} \phi_r \end{cases} \tag{6}$$

The module of rotoric flux is obtained by a block of field weakening given by the following non linear relation:

$$\phi_r^* \begin{cases} \phi_{rN} & \text{if } |\omega| \le \omega_m \\ \phi_{rN} \dfrac{\omega_m}{|\omega_m|} & \text{if } |\omega| \succ \omega_m \end{cases} \tag{7}$$

The rotor flux is controlled by PI controller taking as input the reference value ϕ_r^* and the calculated value.

5. Fuzzy Logic Controller

It appears that fuzzy logic based intelligent control is most appropriate for performance improvement of the ac machines. The main preference of the fuzzy logic is that is easy to implement control that it has the ability of generalisation

[14],[15],[16]. The basic configuration of the fuzzy logic system is featured in Figure 1.

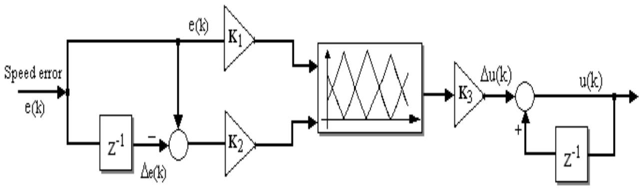

Fig.1 Block diagram of fuzzy control system

In the system presented in this study, Mamdani type of fuzzy logic is used for sped controller. The command signals to the speed controller are the error ‘e(k)’ and change rate of error ‘Δe(k)’. Fuzzy logic controller is based on three well known blocs: Fuzzyfication bloc, block of rule bases and defuzzyfication block, whose function is following briefly explained. The fuzzification stage transforms crisp values from a process into fuzzy sets. The second stage is the fuzzy rule bases which expresses relations between the input fuzzy sets of linguistic description rules A, B and the output fuzzy set C in the form of ″ IF A and B – THEN ″ , and the defuzzyfication stage transforms the fuzzy sets in the output space into crisp control signals. As fuzzy system, we are considering a fuzzy PD controller. The control algorithm is represented by fuzzy rules. The first step in designing the fuzzy controller is to generate the fuzzy rules based on the knowledge of the expert. According to the expert, three situations can be distinguished for the motor speed, namely, above, around and below the desired reference speed. The linguistic representation of the motor speed with respect to a given desired reference speed can be easily translated into a linguistic characterisation of the system error. By defining the system error between the measured speed and the desired speed, the propositions, higher, around and beneath the desired reference speeds are otherwise expressed as Positive, Zero and Negative errors. Furthermore, for given system state variables, the expert can express how he would act if he was controlling the system. For example, a typical rule reads as follows:

IF speed error is Positive Small (**PS**),
AND rate of change in speed error is negative small (**NS**)
THEN change in motor voltage
(Output of fuzzy controller is Zero (**Z**))

The second step consists of modifying the rule-base in order to satisfy the requirements induced by the proposed strategy. The fuzzy controller has to produce a null action when the system has a normal behaviour. In this work, a simple Proportional-Integral type (PI) speed control scheme was implemented and used to assess the basic performance of the system. The output of the fuzzy controller $u_f(k)$ is given by:

$$u_f(k) = F_f\left(e(k) - \Delta e(k)\right) \quad (8)$$

ωhere F_f is a non linear function determined by fuzzy parameters, e(k), Δe(k) are the error and change-of-error respectively. A type of those controllers is fuzzy PI controller whose input is the error e(k).

$$e(k) = \omega_{ref}(k) - \omega_r(k)$$

(9)

where $\omega_r^*(k)$ is the reference model and $\omega(k)$ is the process output at time k. The fuzzy logic controller was used to produce and adaptive control so that the motor speed $\omega_r(k)$ can accurately track the reference command $\omega_r^*(k)$. For the proposed fuzzy controller, the universe of discourse is first partitioned into the five linguistic variables. The controller treats each measurement as a fuzzy singleton and fuzzifies it using the fuzzy sets shown in Figure 2. Triangular shapes were chosen as the membership functions due to the linear equation in evaluation of membership functions. and the output of the fuzzy controller is illustrated in Figure 3.

NB: **N**egative **B**ig
PB: **P**ositive **B**ig
NS: **N**egative **S**mall
PS: **P**ositive **S**mall
ZE: **Z**ero **E**qual.

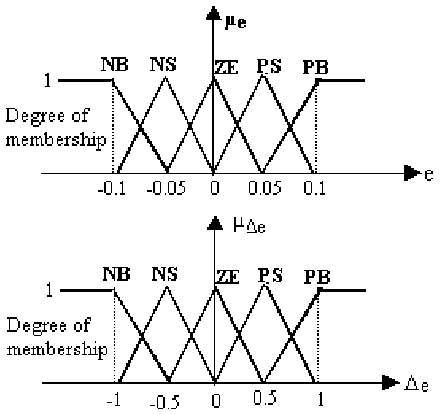

Fig.2 Degree of membership of error and its change

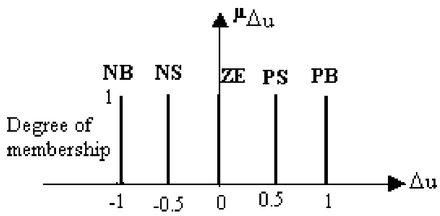

Fig.3 Output membership functions

The fuzzy rules based on speed error e(k) and its variation change $\Delta e(k)$ are presented in Table 1. This implies an inference engine based on 5 implications rules for each of the speed error and its variation, thus a total 25 combinations take place. One can see on Table 1. the rules sets of the fuzzy controller. Every combination is associated to a condition instruction as follows:

If e(k) is **NB** And $\Delta e(k)$ is **PB** , Then $\Delta u(k)$ is **ZE**

TABLE I

Control rules for proposing system

e \ Δ e	NB	NS	ZE	PS	PB
NB	NB	NB	NS	PB	PS
NS	ZE	NS	ZE	PS	ZE
ZE	PB	PB	ZE	PS	NB
PS	ZE	PS	PB	NS	NB
PB	PB	PS	NS	NS	NB

6. MRAS Based For Stator Resistance, Rotor Time Constant and Fuzzy Rotor Speed Estimation

As it is already known, the most difficult aspects concerning the implementation of the electrical drive systems based on the field-orientation theory, are in relation with rotor flux components, speed and resistance parameters estimation. It has been already proved that simultaneous identification of the stator, rotor resistances and the rotor speed is possible only when the rotor flux is time-variant. The overall block diagram of direct field orientation control for induction motor is given in Fig. 4. If the model of the induction motor is considered, the rotor speed and stator resistance and rotor time constant can be identified by approach of Model Reference Adaptive Systems (MRAS) [12]. In the classical MRAS estimation method, there needs two models which outputs are to be compared. One is voltage model (or stator equation) and the other is current model (or rotor equation). The reference rotor flux components obtained from the reference model are given by [13]:

$$\begin{cases} \phi_{r\alpha} = \dfrac{L_r}{L_m}\left(\int\left(v_{s\alpha} - R_s i_{s\alpha}\right)dt - \alpha L_s i_{s\alpha}\right) \\ \phi_{r\beta} = \dfrac{L_r}{L_m}\left(\int\left(v_{s\alpha} - R_s i_{s\beta}\right)dt - \sigma L_s i_{s\beta}\right) \end{cases} \quad (8)$$

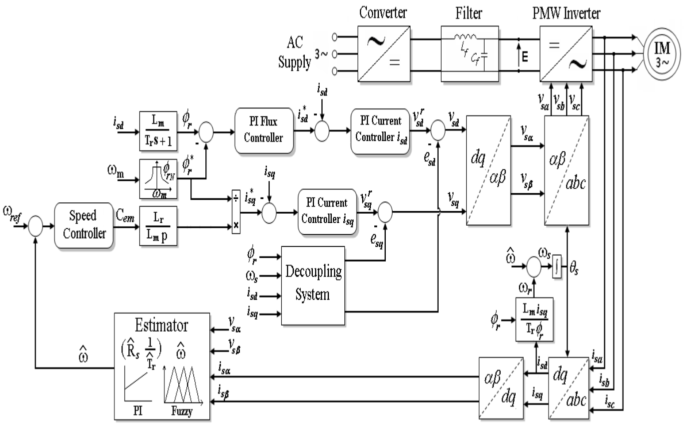

Fig.4 Direct Field-Oriented Control for induction motor equipped with MRAS estimator.

The reference rotor flux components obtained from the reference model are given by:

$$\begin{cases} \dfrac{d\hat{\phi}_{r\alpha}}{dt} = -\dfrac{1}{T_r}\hat{\phi}_{r\alpha} - \hat{\omega}\hat{\phi}_{r\beta} + \dfrac{L_m}{T_r} i_{s\alpha} \\ \dfrac{d\hat{\phi}_{r\beta}}{dt} = -\dfrac{1}{T_r}\hat{\phi}_{r\beta} + \hat{\omega}\hat{\phi}_{r\alpha} + \dfrac{L_m}{T_r} i_{s\beta} \end{cases}$$

(9)

A new structure MRAS is proposed in this paper based to stator resistance, rotor time constant and rotor speed estimation is designed based on the concept of hyperstability, [12],[13] in order to make, the system asymptotically stable. The configuration of the proposed scheme is shown in Fig.5.

The reference model, usually expressed by the voltage model for speed estimation and rotor time constant and current model for stator resistance estimation, represents the stator and rotor equations. It generates the reference value of the rotor flux components in the stationary reference frame from the monitored stator voltage and current components.

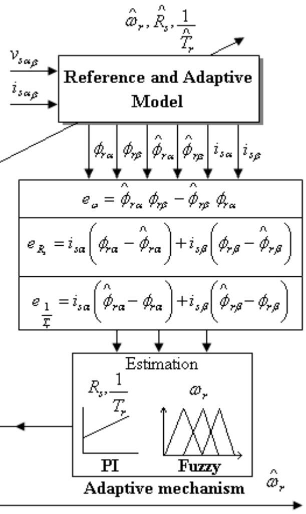

Fig.5 Resistance parameters estimation and rotor speed under MRAS approach

The error equations for the voltage and the current model out puts can then be written as:

$$\begin{cases} e = \underline{\hat{\phi}_r} - \underline{\phi_r} \\ \dfrac{de}{dt} = d\dfrac{\underline{\hat{\phi}_r}}{dt} - d\dfrac{\underline{\phi_r}}{dt} \end{cases}$$

(10)

The equation (10) can be rewritten in matrix notation by:

$$\frac{d\underline{e}}{dt} = A\cdot\underline{e} - W$$

(11)

where
$$A=\begin{bmatrix} -\frac{1}{T_r} & -\omega & 0 & 0 \\ \omega & -\frac{1}{T_r} & 0 & 0 \\ 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \end{bmatrix}$$

(12)

and
$$W=\begin{bmatrix} 0 & -\left(\frac{1}{\hat{T}_r}+\Delta\omega\right) & \frac{\Delta}{T_r} & 0 \\ -\left(\frac{1}{\hat{T}_r}+\Delta\omega\right) & -\frac{1}{T_r} & 0 & \frac{\Delta}{T_r} \\ 0 & 0 & \frac{L_r}{L_m}\Delta R_s & 0 \\ 0 & 0 & 0 & \frac{L_r}{L_m}\Delta R_s \end{bmatrix}\cdot\begin{bmatrix} \hat{\phi}_{r\alpha} \\ \hat{\phi}_{r\beta} \\ i_{s\alpha} \\ i_{s\beta} \end{bmatrix} \quad (13)$$

The adaptation mechanism compares the two models and estimates the speed, rotor time constant and stator resistance by an integral proportional regulator. Using Lyapunov stability theory [10],[11], we can construct a mechanism to adapt the mechanical speed and stator rotor from the asymptotic convergence's condition of the state variables estimation errors. The expressions for the speed and resistance tuning signal and the estimated speed and resistance can be given as :

$$\begin{cases} e_\omega = \hat{\phi}_{r\alpha}\,\phi_{r\beta} - \hat{\phi}_{r\beta}\,\phi_{r\alpha} \\ \hat{\omega} = \left(k_{pw} + \frac{k_{i\omega}}{p}\right)\cdot e_\omega \end{cases}$$

(14)

$$\begin{cases} e_{R_s} = i_{s\alpha}\left(\phi_{r\alpha} - \hat{\phi}_{r\alpha}\right) + i_{s\beta}\left(\phi_{r\beta} - \hat{\phi}_{r\beta}\right) \\ \hat{R}_s = \left(k_{pR_s} + \frac{k_{iR_s}}{p}\right)\cdot e_{R_s} \end{cases}$$

(15)

$$\begin{cases} e_{\frac{1}{T_r}} = i_{s\alpha}\left(\hat{\phi}_{r\alpha} - \phi_{r\alpha}\right) + i_{s\beta}\left(\hat{\phi}_{r\beta} - \phi_{r\beta}\right) \\ \frac{1}{\hat{T}_r} = \left(k_{p\frac{1}{T_r}} + \frac{k_{i\frac{1}{T_r}}}{p}\right)\cdot e_{\frac{1}{T_r}} + \frac{1}{T_r} \end{cases}$$

(16)

K_p and K_i are positive gains.

The fuzzy logic controller of speed estimation is shown in Figures 1. The variables e_ω and Δe_ω presented the input and ω_r the output of fuzzy sensorless. A block diagram of the fuzzy logic sensorless based MRAS structure is presented in Figure 4.

7. Implementation Results and Discussion

The described control structure shows in Figs.4 was implemented in the environment software MATLAB/SIMULINK, and tested in various operating conditions. The numerical method for solving the equations is Runge-Kutta method [19]. Fixed-step mode is chosen for the computational time interval, this will emulate the fixed sampling frequency of the real-time control. The sampling period is 1e-4 sec, the parameters values of the system under study are summarized in Table 2. Fig.6 shows simulation results of rotor speed an external force of 2 N.m, his disturbance can be seen at t = 0.7 sec and t = 1.2 sec and reference change at t = 1.5 s. Fig. 7 shows simulation results of load torque variation.
These results show clearly very satisfactory performance for the proposed sensorless controller in tracking and a remarkable pursuit between measured and estimed

speed of the reference model speed. The control illustrates the correct signal issued by the fuzzy logic controller. There is an excellent direct field orientation consequence of a perfect decoupling between the flux and electromagnetic torque.

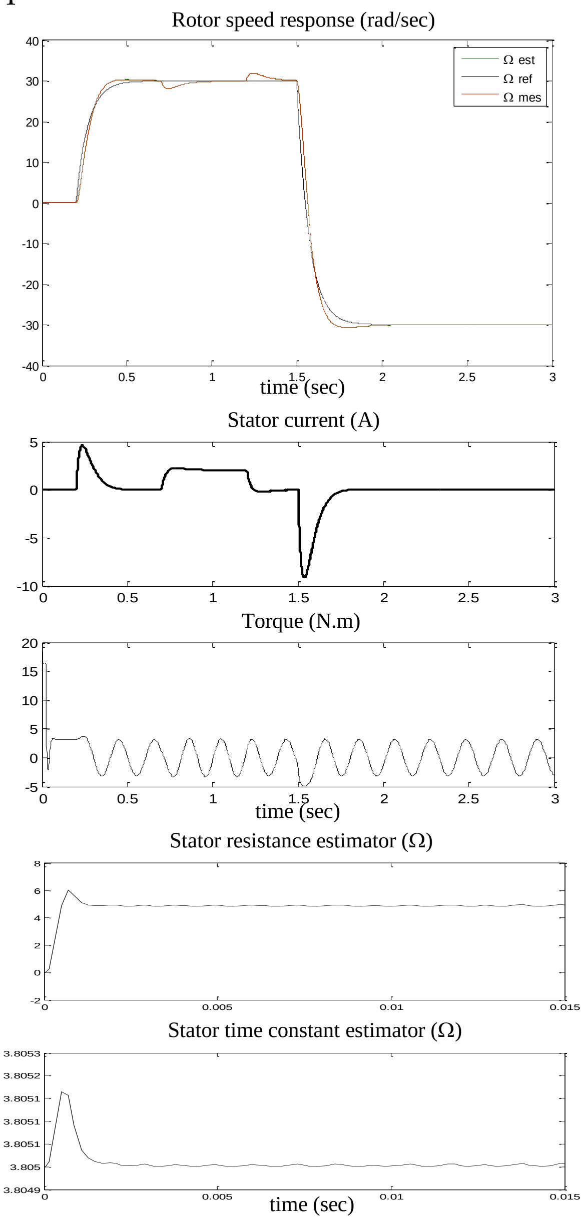

Fig.6 Results of sensorless performance reference and load torque change at t = 0.7 sec and t =1.2 sec and reference change at t = 1.5 sec

Rotor speed response (rad/sec)

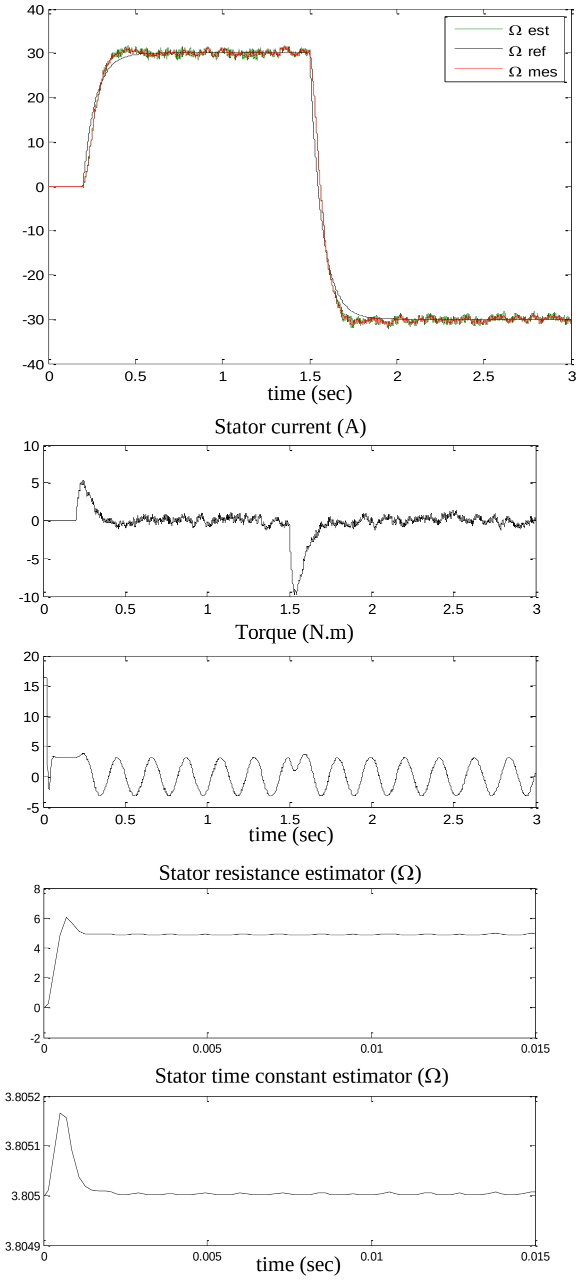

Fig.7 Results of sensorless proposed scheme under load torque variation.

8. Concluding Remarks

A new Fuzzy-MRAS based sensorless technique for vector control of induction motor of algorithm for parallel resistance parameters and rotor speed fuzzy logic estimation applied to direct vector control method is proposed. The system was analyzed and designed. The performance and robustness of the proposed controller scheme have evaluated under a variety of operating conditions of the induction motor drive. The results demonstrated the effectiveness of the proposed structure.

APPENDIX

TABLE II: Rating of tested induction motor

Rated values	*Power*	*1.5*	*kW*
	Frequency	50	Hz
	Voltage Δ/Y	220/380	V
	Current Δ/Y	11.25/6.5	A
	Motor Speed	1420	rpm
	pole pair (p)	2	
Rated parameters	Rs	4.85	Ω
	Rr	3.805	Ω
	Ls	0,274	H
	Lr	0,274	H
	Lm	0,258	H
Constant	J	0,031	kg,m²
	f	0.00114	Kg.m/s

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